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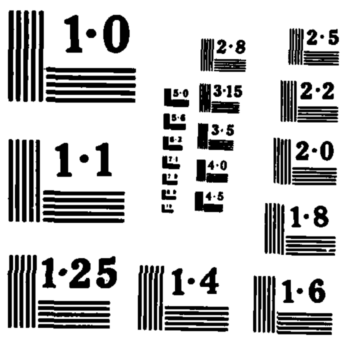

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Final Technical Report
GTRI Project No A-3305

AGILE ELECTROMAGNETICS EXPLOITING HIGH SPEED LOGIC (AEEHSL)

by

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Prepared for

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This document reports research on a 100 MHz, polarization coded, pulse compression, X-band radar. Left circular and right circular polarized signals were generated within each radar pulse to achieve pulse compression with either a coherent carrier or a noise carrier. The radar was designed and built; limited tests were performed.

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PREFACE

The work reported in this final technical report was performed by members of the Radar and Instrumentation Laboratory (RAIL) of the Georgia Tech Research Institute at the Georgia Institute of Technology under Project A-3305. The major contributors from the technical staff of RAIL were:

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R. C. Michelson	Computer Interface Design
D. D. Irwin	Analog-to-Digital Design
J. M. Baden	Software Design
R. D. Sandberg	Software Design
D. W. Sykes	Digital Construction
B. W. Lavers	Digital Board Layout

This project was sponsored by the Department of the Navy under Contract N00014-81-K-0441. The technical monitors for this project were Dr. Ken Davis and Mr. Max Yoder of the Department of the Navy, Office of Naval Research.

EXECUTIVE SUMMARY

This report summarizes the research information that resulted from the program titled Agile Electromagnetics Exploiting High Speed Logic. The key points from this research are:

1. A polarization coded radar was built which performs pulse compression using a polarization code.

2. A Doppler insensitive pulse compression technique was demonstrated.

3. One hundred megahertz digital logic was implemented using conventional construction techniques.

4. Mismatched pulse compression codes were implemented which reduce both the peak and the average range-time sidelobes.

5. Golay code pairs were implemented which yielded zero range-time sidelobes for stationary target radar returns.

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SECTION 1
INTRODUCTION

This final technical report outlines the research performed by the Georgia Tech Research Institute (GTRI) under Navy contract N00014-82-K-0441 entitled "Agile Electromagnetics Exploiting High Speed Logic (AEEHSL)." The objectives of this work were to:

1. Design and construct a radar system capable of modulating the polarization of an X-band carrier frequency at modulation rates up to 100 MHz and
2. Design and construct digital logic circuits capable of 100 MHz speeds.

A measurement radar was designed and constructed with the appropriate polarization modulation characteristics. The radar receiver contains 100 MHz digital logic circuits. The radar is referred to as an Advanced Intrapulse Polarization Agile Radar (A-IPAR). The A-IPAR processing is target Doppler insensitive, it generates a range-polarization profile of the targets within its field of view, it digitizes the 100 MHz bandwidth (or less) radar video, and it performs all post-video signal processing in a slower speed, hardwired digital processor and a microcomputer.

SECTION 2
POLARIZATION CODE THEORY

2.1 POLARIZATION

Radar signals are polarized in the sense that the E-field modulation is constrained to a particular plane in 3-dimensional space or that the plane of modulation can be varied in a particular manner. For example, two orthogonal linear polarizations are vertically (V) polarized and horizontally (H) polarized waves, where the plane of polarization is defined as that perpendicular to which no time dependent E-field modulation can be detected. Electromagnetic waves with elliptical polarizations are generated by simultaneously transmitting sinusoidal H and V polarized waves which differ in amplitude and/or phase. If the amplitudes of the H and V sinusoids are equal and are out of phase by 90°, then the polarization is said to be circular. If the H component lags the V component, the resultant wave's plane of polarization varies according to the right hand rule in the direction of propagation, and the wave is said to be right circularly (RC) polarized. When the H component leads the V component by 90°, the wave is said to be left circularly (LC) polarized. RC polarization and LC polarization are orthogonal.

The circular polarization of a reflected wave is affected by the nature of the reflecting surface. For example, a circularly polarized wave reflection from an odd-bounce scatterer is polarized opposite in sense from the incident wave (i.e., RC incident yields LC reflected and vice versa). As can be deduced from the odd-bounce case, even bounce reflectors leave circular polarization unchanged (i.e., RC incident yields RC reflected). At each bounce, the component of the wave parallel to the surface experiences a 180 degree phase shift, while the component perpendicular to the surface does not experience a phase shift. For the special case of a flat plate and a wave propagation normal to the plate surface, both the H and the V components of the wave experience a 180 degree phase shift. The relative phase between H and V is unchanged, and the direction of propagation is changed by 180

degrees. Therefore, the net result is a change from the incident circular polarization to the circular polarization which is orthogonal to the incident polarization.

2.2 PULSE COMPRESSION

The purpose of pulse compression is to simultaneously achieve high energy in the radiated pulse and fine range resolution. Pulse compression was first achieved utilizing linear frequency modulated (LFM) waveforms. Biphase coding of the RF carrier, as well as discretely stepped interpulse and intrapulse frequency modulation, have also been used to achieve pulse compression. A pulse compression radar codes a transmit pulse of duration T to attain a bandwidth of $1/\tau$, where $T \gg \tau$. By matched filter processing of the echo pulse, the received signal is compressed to a width τ, which achieves a range resolution of 1/2 $c\tau$. A pulse compression system thus achieves the fine range resolution of a pulsed radar that utilizes a pulse of duration τ while in actuality transmitting a pulse of duration T.

2.3 POLARIZATION CODED PULSE COMPRESSION

The A-IPAR transmits a pulse that is encoded by polarization modulation on a subpulse basis. This coding is then utilized to effect pulse compression of the received echo pulse. This method differs from the more classical approaches of pulse compression encoding on carrier phase or frequency in that the coding is contained in the relative phase between the horizontal and vertical polarization components of the transmit pulse.

The A-IPAR system is capable of switching between right and left circular polarizations on an intrapulse basis at rates of up to 100 MHz. Utilizing separate H and V receive ports, A-IPAR downconverts the two channels separately and accurately measures the phase between them. Let ϕ_{HV} represent the phase difference between the H and V channels (that is $\phi_H - \phi_V$) and recall that a signal is LC polarized if $\phi_{HV} = +90°$ and RC polarized if $\phi_{HV} = -90°$. $\sin\phi_{HV} = 1$ for LC signals and $\sin\phi_{HV} = -1$ for RC signals, thus giving a natural correspondence between binary codes and intrapulse polarization

modulation. A-IPAR uses this correspondence to construct well-behaved polarization codes from the class of well-known, well-behaved, binary codes.

Figure 1(A) depicts a natural choice for an IPAR transmit waveform. According to the correspondence described above, the chosen code represents a 13-bit Barker code and yields a pulse compression ratio (T/τ) of 13 to 1. Figures 1(B) and 1(C) represent the expected returns from an idealized flat plate (or any odd-bounce scatterer) and an idealized dihedral (or any even-bounce scatterer).

The A-IPAR processor can compute, on a subpulse basis, $\sin\phi_{HV}$ of any received echo waveform. The results of the computation are passed through a filter matched to the coding on the transmit waveform. The scales for Figure 2(B) apply to both Figures 2(A) and 2(B). Figure 2(A) represents the output from the matched filter compressor given the return, as depicted in Figure 1(B), from an odd-bounce reflector. The time resolution is $\tau/2$ rather than $T/2$ (a 13 to 1 improvement), and the peak signal is T/τ or 13 times the nominal uncompressed level. Figure 2(B) represents the output in response to an even-bounce reflector. Note that the power and resolution characteristics are precisely the same as those of Figure 2(A), although there is a change of sign in signal voltage.

2.4 RELATION TO BIPHASE CODING

Biphase coded signals require a coherent carrier. The A-IPAR technique can use either a coherent or noncoherent carrier, but A-IPAR will not work in some circumstances when using coherent RF signals. The following discussion illustrates this point.

2.4.1 A-IPAR PROCESSING

1. Consider two equal amplitude point targets (T_1 and T_2) spaced by R greater than the range resolution bin ΔR.
2. The round trip phase difference ϕ_T in units of wavelengths between T_1 and T_2 is

$$\phi_T = \frac{2R}{\lambda}$$

where λ is the RF carrier wavelength.

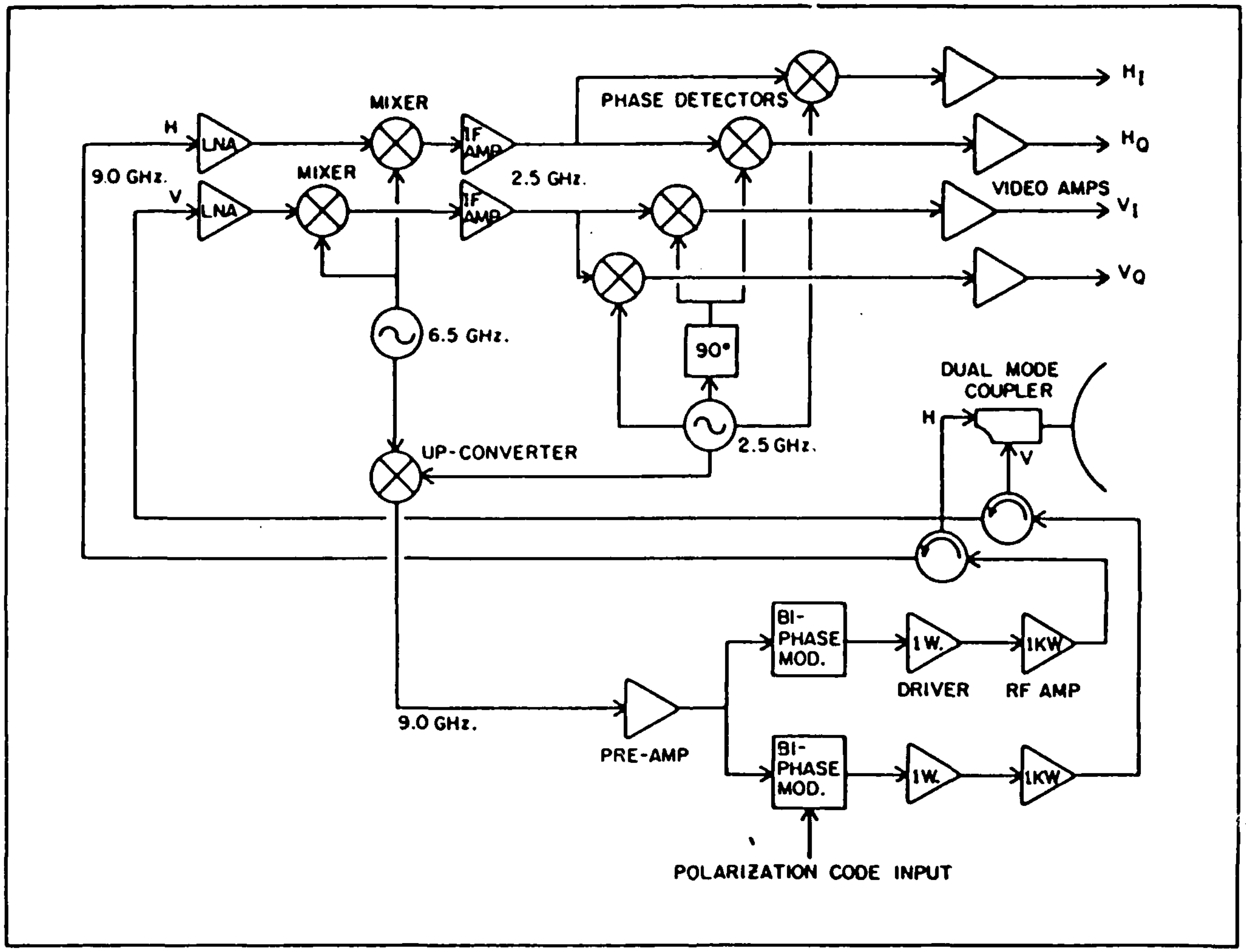

Figure 5. RF subsytem

3.1.1 RF SUBSYSTEM

Figure 5 is a diagram of the RF subsystem. This subsystem includes a two-channel coherent master-oscillator power-amplifier radar transmitter. The two channels are connected to the vertical and horizontal polarization ports of a dual-polarized antenna feed. Whenever a 90^0 phase relationship exists between the two channels, the resultant transmission is circularly polarized. The opposite sense circular polarization is transmitted by switching the phase relationship to a negative 90^0.

The phase shift for one channel referenced to the other is achieved at a low power level due to the speed limitations of high-power phase shifters. For this reason, the A-IPAR system was built with two high-power amplifier channels. The phase shift of one channel is accomplished at low power using a mixer which could conceivably modulate the polarization at rates in excess of 2 GHz for the devices used. The mixers (i.e., phase shifter or phase modulator) have the characteristic that a high-level signal (i.e., +1) at the IF port causes zero phase shift from the input to output, whereas a low-level signal (i.e., -1) causes a phase shift of 180^0. With a permanent 90^0 shift added to the 0^0 and 180^0 phase shift combinations, the result is -90^0 or 90^0. Hence, the required phase relationships are generated for the right and left circular polarization-coded transmissions. The polarization coding is independent of the carrier frequency.

Two receive channels are used with a direct conversion from the 9.0 GHz RF to a 2.5 GHz IF. The 2.5 GHz IF is phase-detected relative to the coherent 2.5 GHz local oscillator. The outputs of the four mixers are the in-phase (I) and quadrature (Q) components of the horizontaly polarized (H) and vertically polarized (V) signals. These four video signals, H_I, H_Q, V_I and V_Q, and V_Q, provide a complete description of the received signal; they contain both the ambiguous total phase of the received signal and the relative phase between the H and V receive channels. These video signals are amplified and routed to the digital processor. Single-channel processing could yield range and Doppler information. Dual-channel (IPAR) processing could yield high-resolution, target-polarization profiles.

TABLE 2. A-IPAR OPERATIONAL PARAMETERS

RF SUBSYSTEM	
Frequency	9.0 GHz
Power	1 kW per polarization channel
PRF's	20-2560 Hz
Pulse Length	0.64-10.24 μs
Subpulse Length	10-160 ns
Minimum Range	50 m
Maximum Range	15 km
IF	2.5 GHz
Phase Matched Channels	Transmit and Receive
DIGITAL SUBSYSTEM	
Maximum Code Length	64 Bits
Received Signal Quantization	6 Bits
Processing	Matched and Mismatched Correlation Filters
Data Recording	Digital Tape
Control	Microprocessor
Range Gate	0 TO 15 km in 0.75 m steps

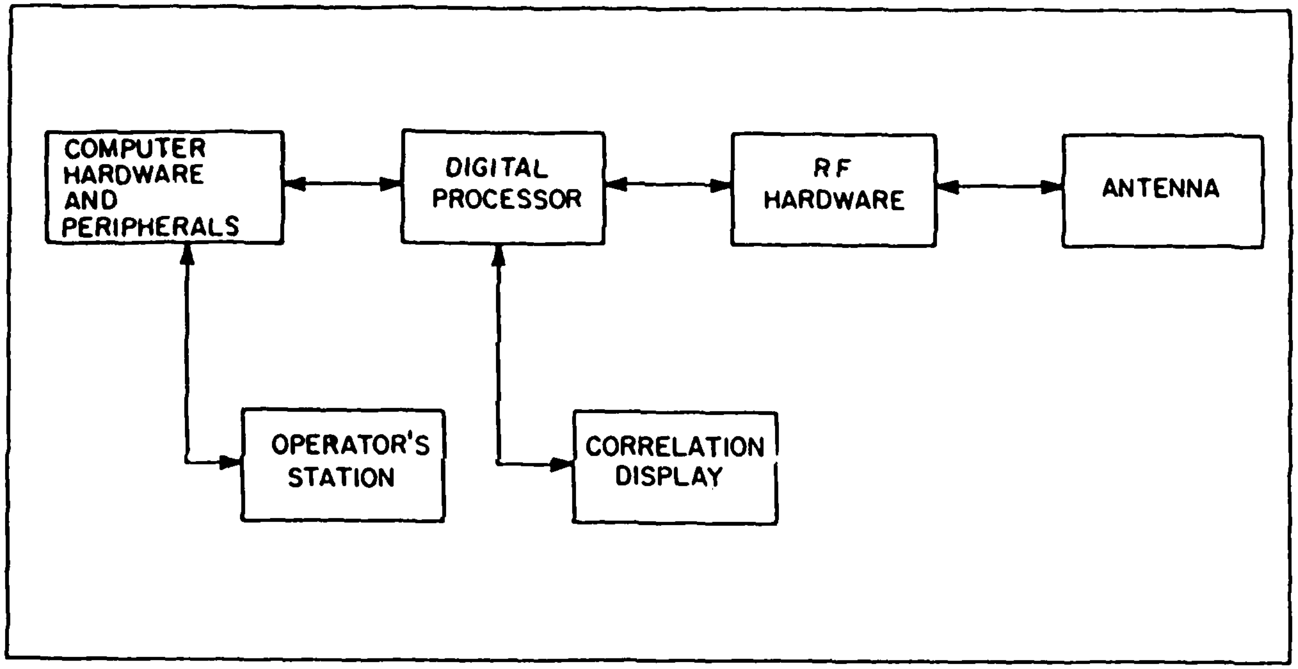

Figure 4. System level block diagram.

SECTION 3
RADAR SYSTEM FUNCTIONAL DESCRIPTION

3.1 SYSTEM LEVEL FUNCTIONAL DESCRIPTION

The A-IPAR System Level Block Diagram is shown in Figure 4. The radar is controlled from an operator's station which consists of a keyboard and a CRT display. The keyboard inputs create a field of parameters which are the control variables of the radar. Parameter selection is aided by a software menu. The selected parameters are displayed on the CRT display. The radar generally responds to the input parameters as they are entered and the operator's choice of parameters is sent by the computer to the digital processor. The digital processor includes five high speed and several low speed circuit boards. The high speed boards (100 MHz) are the High Speed Timing and Control Board, and the four Analog-to-Digital Converter Boards.

The Digital Processor supplies signals to the RF subsystem which enable the polarization code modulation of the electric vector and the pseudo-noise modulation of the phase of the RF carrier. Both these signals are impressed on a low power signal by a mixing process and then amplified to full power for radiation by the antenna.

The energy received by the antenna and receivers in the RF hardware is downconverted to IF and supplied to the Digital Processor. The Digital Processor digitizes four sets of data, which are:

1. In-phase horizontally polarized return (H_I),
2. Quadrature phase horizontally polarized return (H_Q),
3. In-phase vertically polarized return (V_I), and
4. Quadrature phase vertically polarized return (V_Q).

Low speed boards in the digital processor perform correlation, filtering, and display functions for these four sets of data.

The operational parameters for this radar are shown in Table 2.

Summary of Noise Modulated A-IPAR

The analyses of this section show that:

1. The probability of interference between subtargets is a function of the number of subtargets,
2. The degree of interference for a particular set of subtargets may be anywhere in the interval of zero to 100 percent,
3. Wideband noise modulation improves the probability of avoiding interference between targets,
4. If B_N equals the number of subtargets times the code bandwidth, then the probability of code interference is small, and
5. If B_N equals the number of subpulses in the code times the code bandwidth, zero probability of code interference can be achieved.

5. Therefore, the code may be recoverable with recovery percentages between 100 percent and 0 percent.

If the noise bandwidth $B_N > B_C$, the degree of correlation between the paths of the code channel or the paths of the reference channel is decreased. If $B_N \gg B_C$ this correlation approaches zero and the code is fully recoverable. The improvement factor (IF) in dB is given by:

$$IF = 10 \log (B_N/B_C).$$

Table 1 tabulates the IF for different percentage modulation bandwidths at X-band.

TABLE 1. (B_N/B_C) IMPROVEMENT FACTOR

X-Band % BW	RF Noise Bandwidth B_N(MHz)	Code Modulation Bandwidth B_C(MHz)	Bandwidth Ratio (B_N/B_C)	Upper Bound of Improvement Factor in dB
±0.5	100	100	1	0.0
±2.5	500	100	5	7.0
±5.0	1000	100	10	10.0
±10.0	2000	100	20	13.0
±15.0	3000	100	30	14.8
±20.0	4000	100	40	16.0

For the no code modulation case these statements are true:

1. $X'(t)/2 = X(t)/2$
2. $X'(t-\tau)/2 = X(t-\tau)/2$
3. $X'(t)/2 + X'(t-\tau)/2 = X(t)/2 + X(t-\tau)/2$
4. The output of the "H" channel and the "V" channel are correlated.
5. If these two signals are mixed and baseband filtered only DC output will be realized. This result is realized regardless of the bandwidth of $n(t)$. Although $X(t)$ or $X'(t)$ are independent of $X(t-\tau)$ and $X'(t-\tau)$, respectively, the sum shown in step 3 is still correlated.

For a code with a modulation bandwidth $B_C > 1/\tau$ and a noise bandwidth of $B_N >> B_C$:

1. $X'(t)/2$ is independent of $X'(t-\tau)/2$.
2. $X(t)/2$ is independent of $X(t-\tau)/2$.
3. $X'(t)/2$ is not independent of $X(t)/2$.
4. $X'(t-\tau)/2$ is not independent of $X(t-\tau)/2$
5. $[(X'(t)/2 + X'(t-\tau)/2)]$ is not equal to $[X(t)/2 + X(t-\tau)/2]$.
6. The output of the "H" channel and the "V" channel are partially correlated.
7. The output $O(t)$ from the baseband filters is a linear combination of the code bits:

$$O(t) = C(t) + C(t-\tau)$$

8. Therefore, the code is recoverable.

For a code with modulation bandwidth $B_C = 1/\tau$ and a noise bandwidth $B_N < B_C$:

1. The separate paths within the code channel are correlated at delays of τ.
2. The separate paths within the reference channel are correlated at delays of τ.
3. The code channel and the reference channel are partially correlated.
4. The output from the baseband filters is of the form:

$$O(t) = C(t) + C(t-\tau) + C(t)\,C(t-\tau)$$

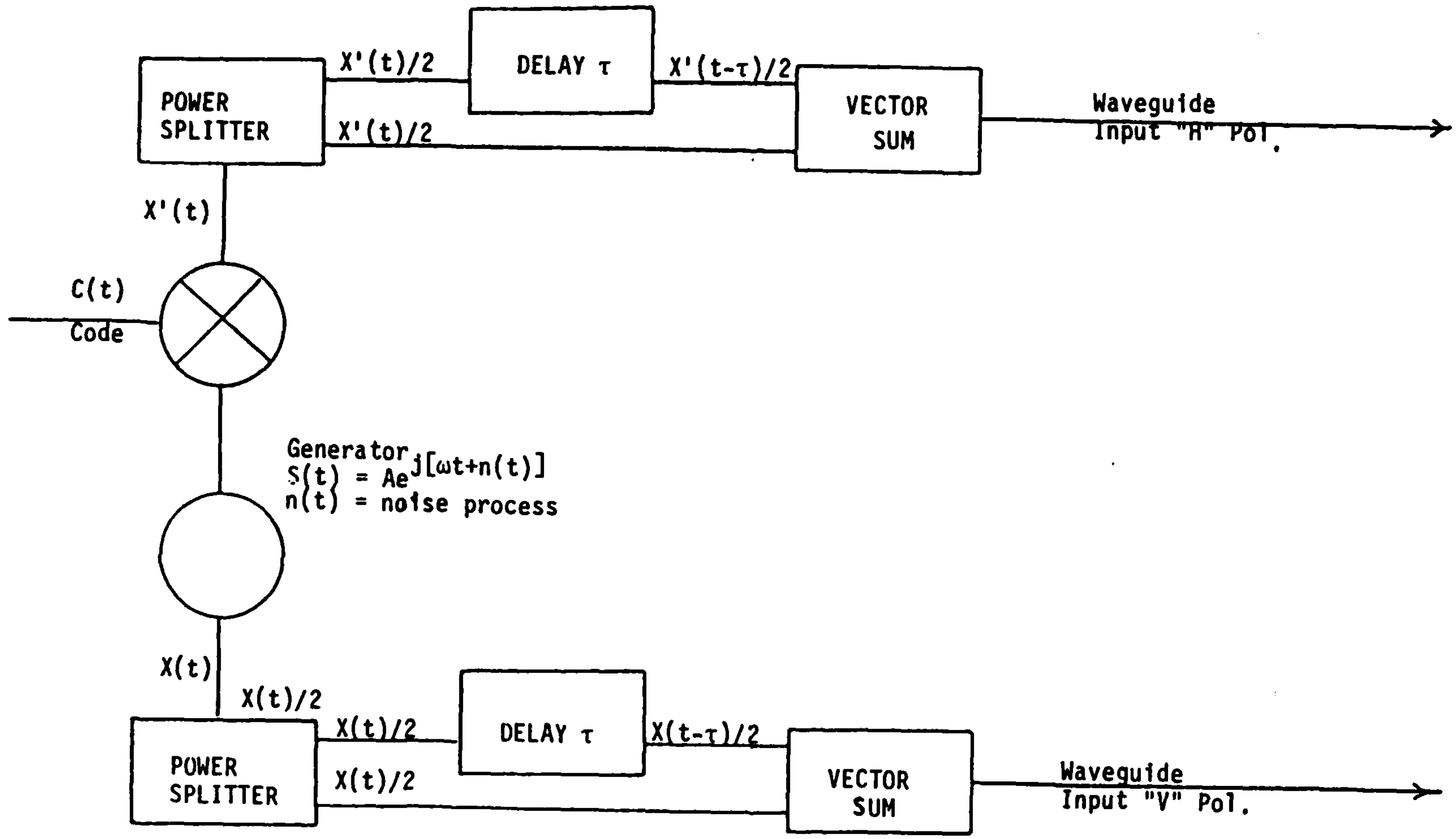

Figure 3. Flow diagram.

2. The biphase coding processor assumes that it experiences the sum of the phase shifts due to relative range and reflections, and it responds correctly to both the relative range and reflection phase shifts.

The key requirement for achieving correct correlation is that the modulation processing scheme must respond correctly to the total phase shift between the target and the radar. A-IPAR meets this requirement for the single point target because the range induced phase shift for each code bit interval is a constant for the single target, and true noise is returned from all code bits except the code bit which is reflected by the target. A-IPAR would also meet this key requirement for extended targets if the RF carrier were noise modulated at a bandwidth which was large relative to the bandwidth of the polarization modulation.

2.4.3 NOISE MODULATED RF CARRIER A-IPAR

Narrowband IPAR signals interact with extended targets to produce signals whose characteristics are not fully compatible with pulse compression. The undesirable/unacceptable signal characteristic is the crosscorrelation between the carrier frequency of signal interval (τ) and signal interval ($\tau + 1$). The purpose of this section is to extend and document the noise modulated RF carrier approach to IPAR processing for extended targets.

Figure 3 is a flow diagram for the "space" portion of the A-IPAR. The interpretation of this diagram is:

1. A generator outputs a signal S(t) composed of a carrier phase (ωt) and a noise phase n(t).
2. A code modulator (mixer) is available to code modulate S(t) for the horizontal polarization channel.
3. The unmodulated signal X(t) and the code modulated signal X'(t) are both power split.
4. One path of "H" and one path of "V" is time delayed by τ.
5. The time delayed signals are vector summed with the undelayed signals.
6. The output from the vector sum is equivalent to the return at the radar feed for an M subpulse code from an N = 2 subtarget target with M > N.

2.4.2 BIPHASE PROCESSING

1. Consider two equal amplitude point targets (T_1 and T_2) spaced by R greater than the range resolution bin ΔR.

2. For these two targets, the round trip phase difference is

$$\phi_T = \frac{2R}{\lambda}.$$

3. Because R is arbitrary, ϕ_T will be a selection from a uniform distribution over the interval 0 to 360 degrees.

4. In space, the electric field vectors from T_1 to T_2 add at a relative phase between 0 and 180 degrees.

5. Let a two-pulse code of the from (+1,+1) be reflected such that T_1 equals T_2 except for a 180 degree phase shift.

6. If T_1 and T_2 were measured separately, then the assigned codes would be $T_1 = +1$ and $T_2 = -1$.

7. The correlation function from these separately measured signals would be zero.

8. The vector addition in the feed yields zero voltage, which yields a zero correlation.

These discussions of A-IPAR and biphase signal modulation and processing reveal that A-IPAR does not work correctly for a coherent carrier and range-extended targets. In contrast, biphase coding does work correctly for the coherent carrier and range-extended targets. The differences are:

1. The A-IPAR processor assumes that only the relative phase shifts due to reflection are experienced, but it actually responds to both the reflection phase shift and the phase shifts due to relative range.

3. Since R is arbitrary, ϕ_T will be a selection from a uniform distribution over the interval 0 to 360 degrees.

4. The electric field vectors of the signals reflected from T_1 and T_2 will have a relative phase difference uniformly distributed on the interval 0 to 180 degrees.

5. Let a two-pulse code of the form (+1,+1) illuminate these two targets, and let the two returns be present at the radar antenna at the same time.

6. Then the magnitude of the composite electric field vector at the radar can vary from zero to twice the magnitude of the individual electric field vectors.

7. If each target reflected its component of the (+1,+1) code as (+1,+1), then the result from the A-IPAR correlator should be +2.

8. Because of the vector addition in space (i.e., at the antenna), the correlation output is uniformly distributed from 0 to 2.

9. For this special coherent RF case the correlation should be +2, but the actual value of the correlation is a selection from a uniform distribution bounded by 0 and 2.

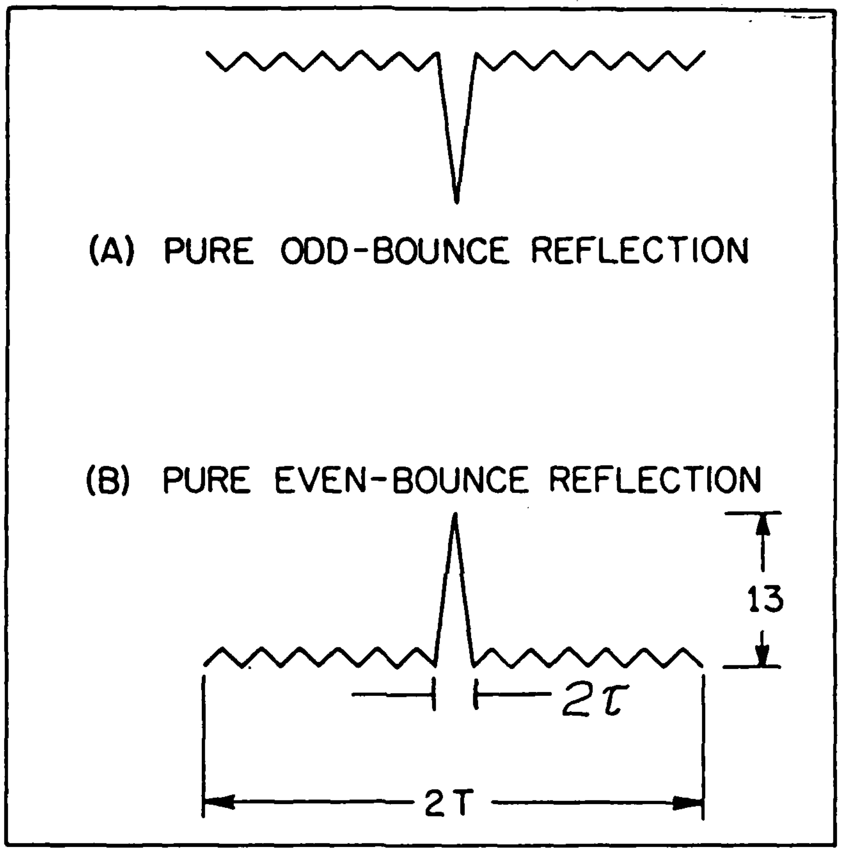

Figure 2. Matched filter compressor output.

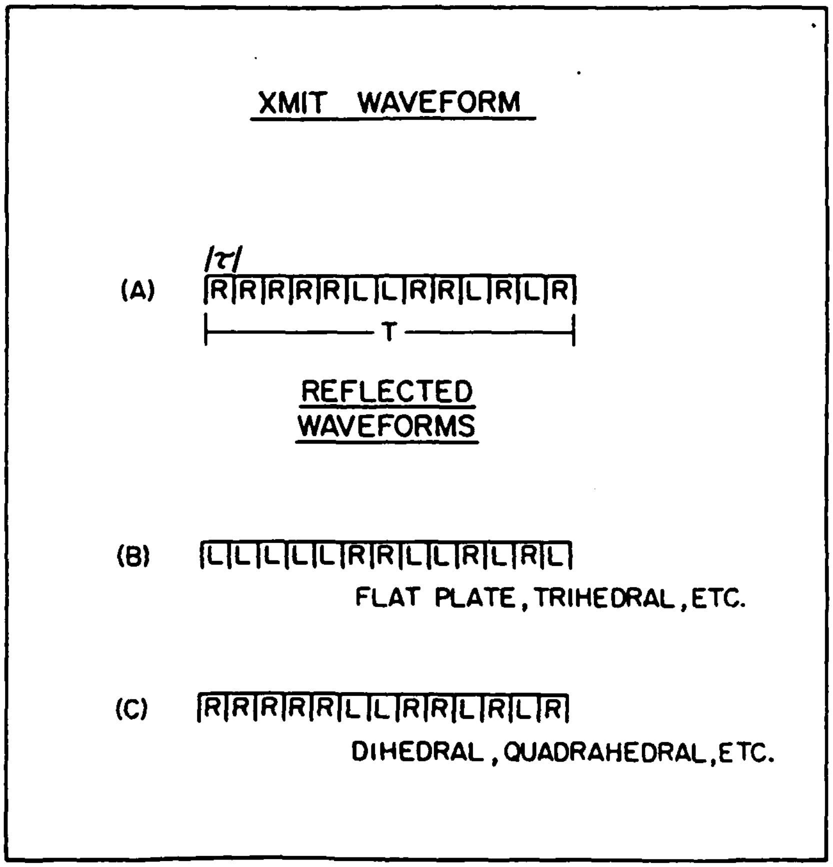

Figure 1. Transmitted and reflected waveforms.

3.1.2 DIGITAL SUBSYSTEM

The diagram for the A-IPAR digital processor is shown in Figure 6. The four video signals from the RF Subsystem (H_I, H_Q, V_I, V_Q) are amplified and applied to 100 MHz 6-bit quantization A/D converters, with temporary digital data storage on the A/D circuit board. This combination of 100 MHz A/D converters and digital data storage allows a data-rate conversion from the input 100 MHz to the PRF of the radar.

THe A/D outputs are routed to both the signal combiner and a digital tape recorder interface circuit board. The signal combiner converts the four components of the two polarization channels into the appropriate scalar form for correlation processing. The correlated outputs (pulse-compressed signals) are routed through a digital filter and a D/A converter. The output of the D/A converter is displayed on an oscilloscope.

The basic control of the A-IPAR system is divided between the hardware digital processor and the microprocessor. The microprocessor establishes the operational mode of the A-IPAR system, while the hardware digital processor establishes both the low speed and the high speed timing and control used during the execution of this mode.

3.1.3 COMPUTER HARDWARE AND SOFTWARE

The A-IPAR uses an AIM-65 computer for mode control and data initilization. The AIM-65 functional relationships to the A-IPAR system are shown in Figure 7.

The three available AIM-65 modes as illustrated in Figure 8 are:

1. RUN Mode - This mode controls the radar system parameters and the data recorder.

2. CODE Mode - This mode enables the examination and alteration of codes and filter weights

3. READ Mode - This mode enables the reading or replaying of the data from the digital tape recorder.

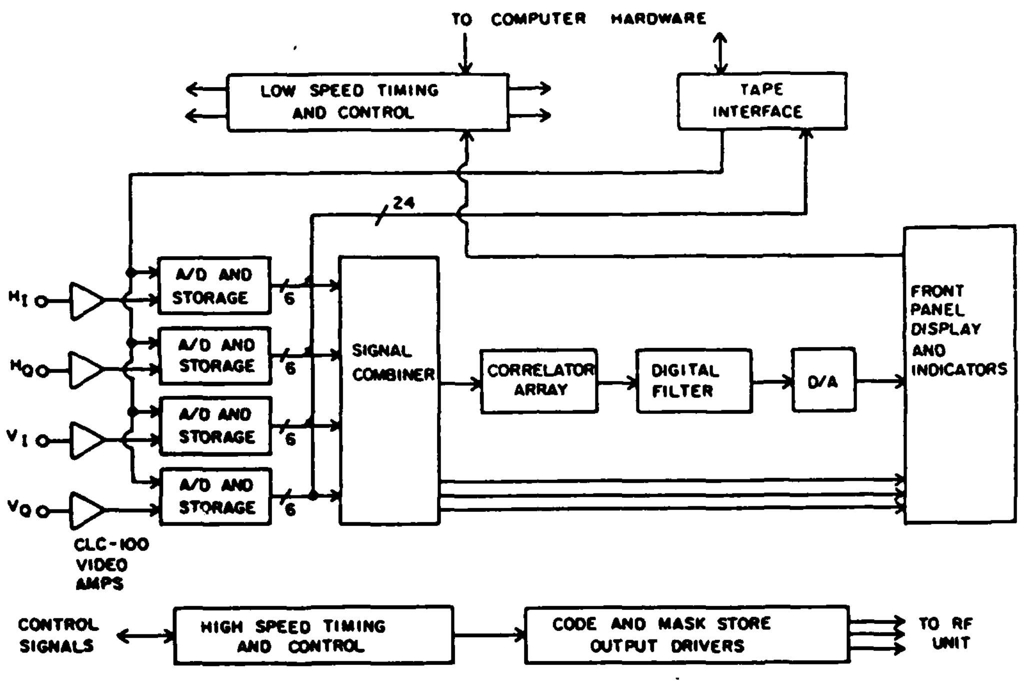

Figure 6. Digital subsystem.

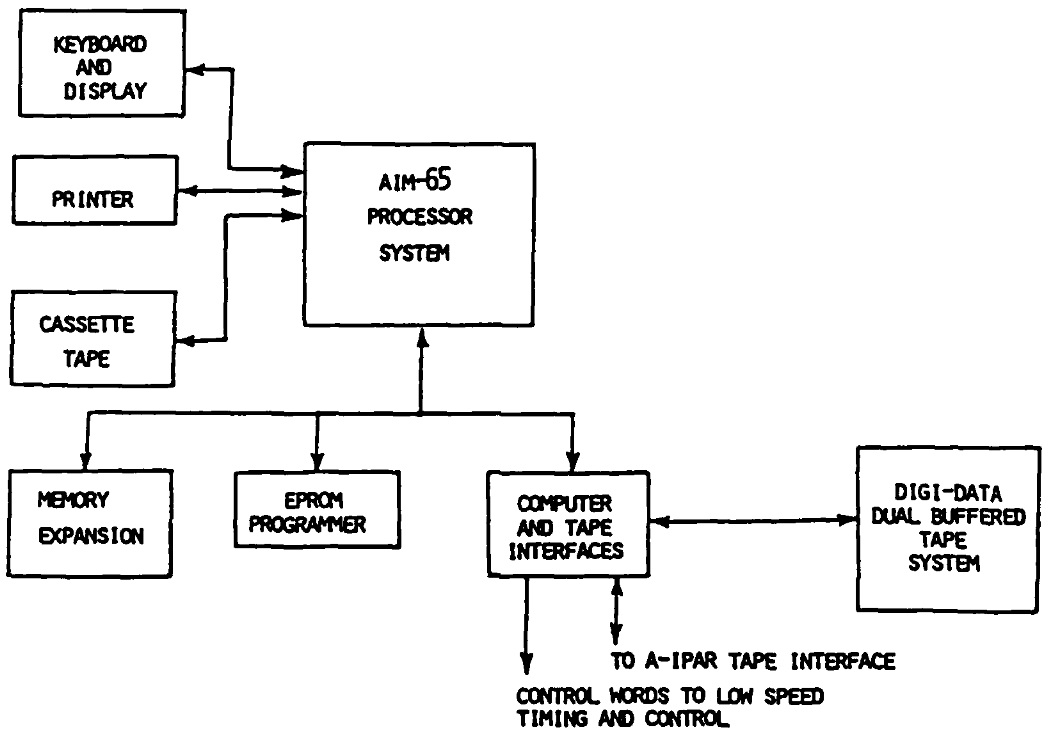

Figure 7. Computer hardware.

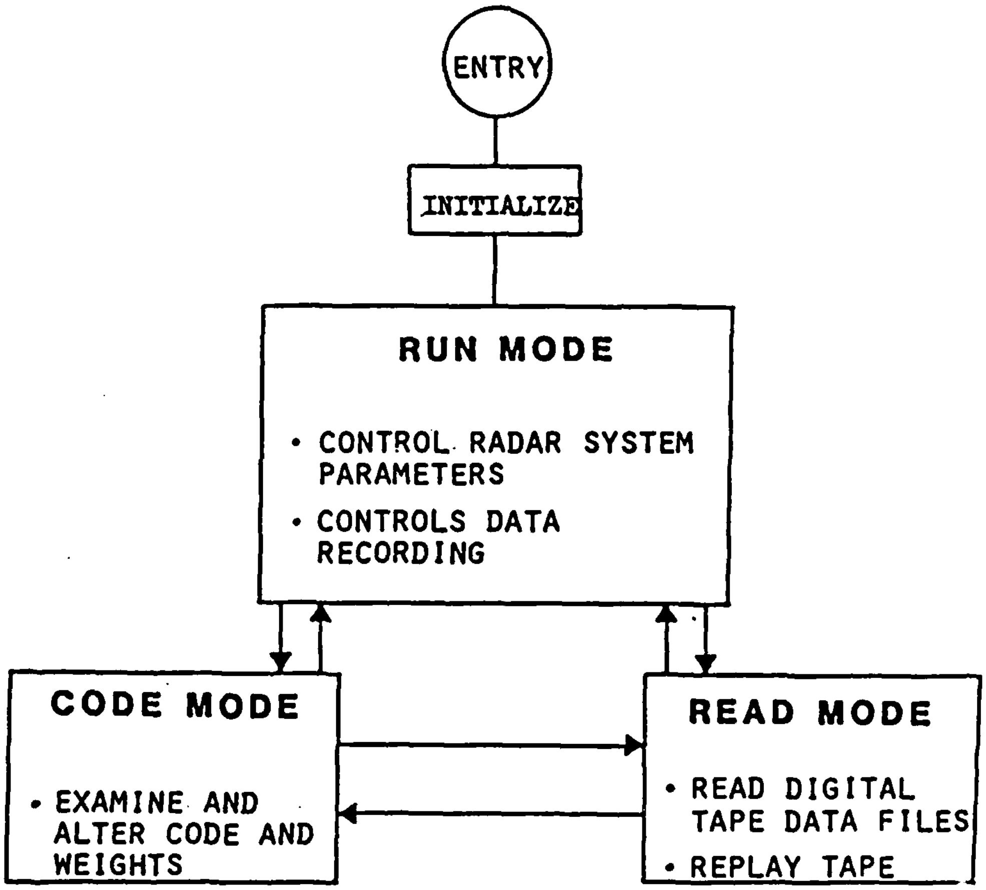

Figure 8. Software modes.

The RUN mode enables the high speed and timing and control subsystem. The clock signals available in this subsystems are used to initialize the radar, clock the code from the high-speed code storage memory to drive the code modulator, delay the data window for some time period ΔT after the signal transmission, and clock the A/D converter boards.

The high speed timing and control signals control the low speed timing and control subsystem which supplies timing and control signals to the tape recorder, the correlator, the digital filter, the D/A converter, and the display oscilloscope.

The CODE mode is used to change the transmitted codes and the filter weights applied to the correlation process. There is storage space within the high speed memory for 32 codes of length 64 bits or less. The radiated code can be changed by a keyboard command which specifies the address of the code. There is a provision for one specified code address to alternate between two specific codes on a pulse by pulse basis. A standard code library is stored in a low speed memory (PROM). This library currently consists of two code sets. Each code set contains sixteen codes and associated filters. One of these code sets is shown in Table 3. Either of the two code sets may be loaded into the RAM on the high speed board while in the CODE mode.

The READ mode allows tape recorded data to be read and played back through the low speed processor portion of the system.

TABLE 3. PROM CODE LIBRARY

CODE #	CODE LENGTH, TYPE	Filter	Performance (in dB) PSL*	ISL*	LPG*
1	64 Uniform	Matched	-.1	+16.2	--
2	16 Alternating	Matched	-.6	+9.9	--
3	64 Alternating	Matched	-.1	+16.2	--
4	5 Barker	Matched	-14.0	-7.0	--
5	5 Barker	Mismatched, 19 Bits	-35.3	-25.3	0.6
6	13 Barker	Matched	-22.3	-11.5	--
7	13 Barker	Mismatched, 35 Bits	-37.7	-24.9	0.3
8	25 Concat. Barker (5 x 5)	Matched	-14.0	-5.0	--
9	25 Concat. Barker (5 x 5)	Mismatched, 53 Bits	-26.9	-18.1	1.0
10	33 Concat. Barker (11 x 3)	Mismatched, 35 Bits	-19.1	-11.3	1.7
11	39 Concat. Barker (11 x 3)	Mismatched, 39 Bits	-27.2	-17.8	1.4
12	52 Concat. Barker (13 x 4)	Matched	-12.0	-5.9	--
13	52 Concat. Barker (13 x 4)	Mismatched, 52 Bits	-25.0	-16.5	1.8
14	64 Concat. Barker (13 x 5-1)	Mismatched, 64 Bits	-25.0	-13.8	0.7
15	64 SLC Pair	Matched	-17.0	-4.8	--
16	64 SLC Pair	Matched	-17.0	-4.8	--

*PSL = Peak Sidelobe Level = $\frac{\text{Largest Sidelobe}}{\text{Signal Peak}}$ in dB

ISL = Integrated Sidelobe Level = $\frac{\text{Total Energy in Sidelobes}}{\text{Signal Peak}}$ in dB

LPG = Loss in Processing Gain = $\frac{\text{Matched Filter Signal Peak}}{\text{Mismatched Filter Signal Peak}}$ in dB

SECTION 4
SUMMARY OF ALL WORK ACCOMPLISHED

4.1 ANALYSES

Under the A-IPAR project the following analyses were performed.

No.	Subject	Result
1	Correlation Code Theory	Matched and mismatched codes were designed which reduce the range time sidelobes. A Golay code pair was selected which cancels the range time sidelobes.
2	Polarization Code Theory	Polarization coded pulse compression theory was extended to scatterers in multiple range bins with coherent and noncoherent carriers.
3	Scattering Matrix Data	Scattering matrix data collection methods on a single PRI or two PRIs were defined.
4	Target Texture	Preliminary investigations were undertaken of radars which are sensitive to the roughness of reflecting surfaces.

4.2 DESIGN

Under the A-IPAR project the following designs were developed.

No.	Subject	Result
1	Van Electrical and Mechanical	A van design was developed to receive the A-IPAR cabinets.
2	Rack Electrical and Mechanical	A design was developed for three racks to contain the A-IPAR hardware.
3	RF	An RF design was developed with a bandwidth specification of 500 MHz.
4	Digital	A digital design was developed with a bandwidth specification of 100 MHz.
5	Software	A software design was developed.

Demanding problems which were solved included the design of the 6 bit, 100 MHz analog-to-digital converter boards and the design of a correlator which could process matched, mismatched, and Golay codes.

4.3 CONSTRUCTION

Under the A-IPAR project these constructions were completed:

No.	Subject	Result
1	Electrical and Mechanical for Van and Racks	Functional equipment

No.	Subject	Result
2	RF	Functional equipment at 100 MHz bandwidth.
3	Digital	Functional equipment at 100 MHz bandwidths and 6 bits of amplitude.
4	Software	Functional software

4.4 TEST

Under the A-IPAR project limited demonstration tests were performed. A definite need exists to extend these tests under a systematic plan for qualifying the A-IPAR system and for quantifying A-IPAR's measurement characteristics.

4.5 GOALS

The A-IPAR project met many but not all of its original goals. Specifically, the test goal for the A-IPAR and the 500 MHz digital hardware goal was not achieved. Nevertheless, the A-IPAR project was a successful effort in that it advanced the central goals of this type of research:

> Investigation and demonstration of the joint interaction between high speed digital hardware and radar system concepts.

SECTION 5

INDEX OF ALL TECHNICAL REPORTS

Under the A-IPAR project these technical reports were presented:

NO.	TYPE	NAME	PLACE
1.	Briefings	Three Technical Reviews of Project	ONR* - GTRI - NRL
2.	Letter Reports	First yearly Letter Report (July 1983)	ONR
3.	Letter Report	Second Yearly Letter Report (July 1984)	ONR
4.	Internal Technical Memorandum	Pulse Compression Coding Study (12 January 1983)	GTRI (Extracted into ONR & NAVSEA Briefings)
5.	Final Technical Report	Third yearly Report (July 1985)	ONR

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SECTION 6
INDEX OF PUBLICATIONS

In association with the A-IPAR project, these publications were generated.

NO.	TYPE	NAME	PLACE
1.	Magazine Article	Advanced Intrapulse Polarization Agile Radar Nears Deployment	Microwave System News February 1984 (Page 67)
2.	Conference Paper	Optimal Mismatched Filters for Pulse Compression Radars	IASTED Paris, France June 1985 (To be published in Conf. Proc.)